A solution for Equalization in Wireless Channel Communications

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Abstract: - The novelty of the solution consists in a blind equalization/detection (BED) algorithm that replaces the classical differential encoding/decoding (DED) equalization technique. In this context, blind means that no pilot carriers or training blocks are required in order to perform channel estimation. Our new BED approach utilizes the sequence detection technique for joint equalization and decoding of OFDM signals over frequency-selective channels, based on the fact that the inherent Fourier transform processing turns a single wideband frequency-selective channel into a set of narrowband frequency flat fading channels. The proposed detection strategy was implemented on a TMS 320 C31 DSP via a standard Viterbi algorithm. In the proposed detector, channel equalization is accomplished by means of a trellis detector operating on a block-by-block basis. We have defined the resulting scheme a blind Viterbi equalizer (BVE). The simulation results show that BED performs better than DED at high signal-to-noise ratio and does not require any periodic training information. Also, we show that BED operates well in a time-varying environment and so it is an attractive solution for high speed multimedia application.

Key-Words: - Coded orthogonal time division multiplexing, Channel estimation, Blind equalization/detection, Viterbi detection, Multicarrier modulation.

1 Introduction

Broadband wireless technology has an important role in the evolution of future global communications. Multicarrier modulation is an attractive technique to overcome the delay spread problem. In such transmissions, the high-rate serial data stream is split into many low-rate parallel streams, modulating orthogonal carriers in subjacent subbands. For each subcarrier, a frequency-flat channel model can be assumed, and consequently the channel distortion can be compensated if an estimate of the channel frequency response is available. With this aim, standard OFDM solutions envisage the periodic insertion of training data blocks to perform channel estimation, leading to the use of Differential Encoding/Detection (DED), but this approach is reliable only when the variations of the channel frequency response between adjacent subcarriers are negligible. In this paper, we use another approach, defined as Blind Equalization and Detection. In this context, blind means that no pilot carriers or training blocks are required in order to perform channel estimation. For this approach, we consider an idea proposed in [1] and [2], that sequence detection techniques used for time–selective flat fading channels are suitable to be applied for joint equalization and decoding of OFDM signals over frequency-selective channels. In the proposed solution, channel equalization is accomplished by means of a trellis decoder, while the equalization algorithm exploits the frequency correlation between subcarriers to estimate the channel frequency response.

2 The description of the transmission system

2.1 The transmitter

A serial stream of M-PSK data symbols with the bit rate of $1/T_s$ is partitioned by a serial-to-parallel converter into blocks of length N, forming a sequence $a_N = [a_0, a_1, a_2, ..., a_{N-1}]$ of frequency-domain symbols. This sequence is input to an Inverse Discrete Fourier Transform (IDFT) that produces the N-dimensional block $b_N = [b_1, b_2, ..., b_{N-1}]$ time-domain symbols.

After the IDFT is computed, a cyclic prefix of length N_1 and a cyclic postfix of length N_2 are appended to the vector b_N in order to form the cyclically extended block of overall length $N_T = N_1 + N + N_2$

$c_N = [c_{-N1}, ..., c_{1}, c_{0,....}, c_{l}, ..., c_{N}, ..., c_{N+N2-1}]$.

Finally, the cyclic extended block feed the linear modulator with impulse response $p(t)$ windowed to an overall length of $L_p << N$ symbol intervals, such as the baseband equivalent signal at the modulator output is:

$$s(t) = \sum_{l=-\infty}^{+\infty} \sum_{k=-N_1}^{N+N_2-1} c_k p(t - kT_s - lN_T T_s) \qquad (1)$$

The waveform (1) is transmitted over the multipath channel, assumed to be stationary, with the impulse response function $h(t)$ either deterministic or a

Gaussian process. In such conditions, the received signal $r(t)$ can be expressed as:

$$r(t) = \sum_{l=-\infty}^{+\infty} \sum_{k=-N_1}^{N+N_2-1} c_k g(t - kT_s - lN_T T_s) + n(t) \qquad (2)$$

where $n(t)$ is a White Gaussian Noise (WGN) with the spectral density $2N_0T_s/E_s$, E_s being the transmitted energy per symbol and :

$$g(t) = p(t) \otimes h(t) \qquad (3)$$

Fig.1 depicts the main blocks of the transmitter structure.

is the noise at the filter output. We have considered that the length of $f_l(t)$ does not exceed $2L_p+L_d$ symbol intervals, where L_d is the maximum channel delay spread.

In the time domain, intersymbol interference (ISI) appears both as interblock interference (due to symbols belonging to different blocks, i.e. in terms with $l \neq m$ in (4)) and as intrablock interference (due to symbols belonging to the same block, i.e. in terms with $l=m$ and $i \neq m$ in (4)). These effects can be minimized or even cancelled through the addition of the prefix and postfix, if the condition $N_1+N_2 \geq 2L_p+L_d$. is respected, the timing error is turned into a linear phase shift at the DFT output. This additional phase shift will be compensated by the BED algorithm described in the next section.

The input in the S/P block for the m-th symbol block, after the removal of the prefix and the postfix is:

$$x_m[n] = \sum_{i=-N_1}^{N+N_2-1} c_i^{(m)} f_m\big((n-i)T_S + \varepsilon_m T_S\big) + w_m[n], \qquad (5)$$

$$n = 0,1,\ldots,N-1$$

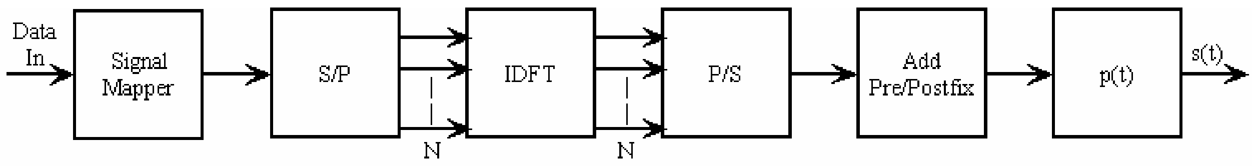

Fig. 1. Block scheme of the transmitter

2.1 The receiver

The first processing block in the receiver is the matched filter. Assuming that the m-th block of length l is to be processed, the matched-filtered received signal is sampled at the instants $t_{n,m}=(mN_T+n)T_s+\tau_m$, where $n=-N_1,\ldots,N+N_2-1$ and τ_m is the m-th local estimate of the overall channel propagation delay τ. From (2) the n-th sample of the m-th block at the matched filter output is :

$$x_m[n] = \sum_{l=-\infty}^{+\infty} \sum_{i=-N_1}^{N+N_2-1} c_i^{(l)} f_l\big(t_{n-i,m-l} + \varepsilon_m T_S\big) + w_m[n], \qquad (4)$$

$$n = -N_1,\ldots,N+N_2-1$$

where $f_l(t) = g(t) \otimes [p(-t)/T_s]$, $\varepsilon_m = (\tau_m - \tau)/T_s$ and $w_m[n]$

We can now consider that after the computing of DFT the output sequence is:

$X[k]=a_kF_\varepsilon[k]+W[k]$, $k=0,1,\ldots N-1$ (6)

where $F_\varepsilon[k]$ are samples of the Fourier transform of the time sequence $f_\varepsilon[n]$ and $W[k]$ is the Gaussian vector corresponding to $w[n]$. If we consider the negative frequencies too, equation (6) becomes:

$X[|k|_N]=a_{|k|_N}F_\varepsilon[k]+W[|k|_N]$ $\quad -N_\alpha \leq k \leq N_\alpha$ (7)

Fig.2 depicts the main blocks of the receiver structure.

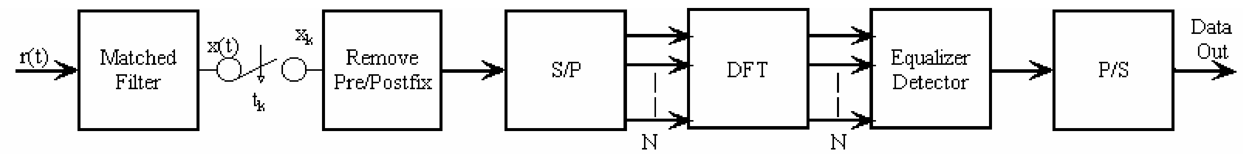

Fig. 2 Block scheme of the receiver

In [3], it is shown that if we consider that the bandwidth of f(t) is $B=(1+\alpha)/(2T_s)$, the synchronization error ε introduces at the DFT output a linearly varying phase shift in the samples with $0\leq k\leq N_\alpha$ where $N_\alpha=int[N(1-\alpha)/2]$. The suggestion is to suppress N_{sc} subcarriers, $N_{sc}=N-1-N_\alpha$ in the roll-off region designed by the value of α, but of course this means a loss in the transmission efficiency.

3 A new algorithm for joint blind equalization and detection

We assume that the transmitted data block $a_N=[a_0,a_1,a_2,...,a_{N-1}]$ represents the differential encoding of the M-PSK source data block $d_{N-1}=[d_0,d_1,d_2,...,d_{N-1}]$ according to the encoding law:

$$a_{|k|N} = a_{|k-1|N}\, d_{|k|N}$$

with $k=-N_\alpha+1,....,N_\alpha$ and $a_{N-N\alpha}=1$, i.e. applies to the nonzero elements of d_{N-1} since the N-$2N_\alpha$-1 central inputs are set to zero to suppress the corresponding carriers.

We propose now the following procedure:

- the ML estimate of d_{N-1} can be expressed [4] as:

$$\bar{d}_{N-1} = \arg\min \sum_{k==n_\alpha+1}^{N_\alpha} \left\{ \left| X\left[|k|_N\right] - \bar{a}_{|k|N}\bar{F}_\varepsilon[k] \right|^2 \right\} \quad (8)$$

The expression between {} will be denoted as $\Gamma(X[|k|_N)], \bar{F}\varepsilon[k], \bar{a}_{|k|N})$

- $F\varepsilon[k]$ may be estimated by using L consecutive measurements of: $Y(k,a_{|k|N}) = X[|k|_N]\ \bar{a}_{|k|N}$ i.e. $\{Y(k-i,a_{|k-i|N}),\ i=1,...,L\}$. For this purpose, we

$$\bar{F}_\varepsilon(f) = \sum_{i=0}^{P} \alpha_i \left[(f-f_0)NT_s\right]^i$$

model $\bar{F}\varepsilon[f]$ as a polynomial of order P in the frequency interval $f\in[(k-1)/NT_s,(k-L)/NT_s]$, where f_0 is a reference frequency and we look for the coefficients $\{\alpha_i, i=0,1,...P\}$ yielding the minimum mean-square-error between the measurements: $Y(k,a_{|k|N})$ and the predicted values $\bar{F}\varepsilon(k/NTs)$, which can be considered as an estimate of $\bar{F}\varepsilon(k)$.

- At a low SNR, the dispersion of the noise increases with P and decreases with L, but at high SNR it is preferable to increase P at a given L [5]. In our simulations, we tried to adequate to various applications the selection of P and L, but the significant results was obtained with P=2 and L=3 and 4.
- In order to obtain a blind detection algorithm that makes the joint between channel equalization and data detection we employ the quantities of $\Gamma(X[|k|_N)], \bar{F}\varepsilon[k], \bar{a}_{|k|N})$ as branch metrics in the same decoding trellis adopted for ML detection. In conclusion, the trellis states at the k-th step are considered as any possible vector of R past symbols: $\delta_k = (\bar{a}_{|k-1|N}, a_{|k-2|N}, ..., a_{|k-R|N})$.

In the case of the M-PSK modulation, we have M branches starting from a state δ_k and going to different states δ_{k+1}. For each trellis state, a channel parameter is estimated from one of the prediction equation generated with a pair (P, L); then, we can compute the branch metrics with the relation (8) and finally we can use them in a Viterbi algorithm.

4 Implementation

4.1 Specific details

For the purpose of method illustration and comparison, the following details are assumed:

- *Transmission Format*: QPSK is assumed with 25% roll-off at both transmitter and receiver. The postfix and prefix are $N_1 = N_2$ =150 symbols intervals; the data packet contains 800 QPSK symbols without overhead.
- *Channel Model*: The channel is assumed with N-path of equal power, uniformly spaced by the interval (symbol period) T=1 μs. Two scenarios can be investigated, with the channel assumed as static and with the channel assumed as quasi-stationary. In both cases mutually independent Gaussian N=2 and N=6 paths can be assumed, with an exponential power delay profile. The average delay is considered $5T$ and the maximum delay as $20T$.
- *Coding Scheme*: A rate ½ convolutional code with a constraint length of six is used. The generator polynomials are $g_1(x)=x^5+x^3+x+1$ and $g_2(x)=x^5+x^4+x^3+x^2+1$; this code ensures the minimum Hamming distance of 7. For decoding a 32-state Viterbi decoder is employed.
- *The symbol rate*: According to the symbol period, the symbol rate is 2 Mbaud; this entails that L=40 to cover the maximum delay.
- *OFDM subcarriers*: In the simulations, we take N=1024 OFDM subcarriers; the number of suppressed carriers is consequently N_{sc}=256.
- *Antenna Diversity:* Two-branch diversity was simulated by using a delay of one symbol unit between both antennas to create an artificial multipath effect; as an example, flat fading can be introduced on a branch.

In order to allow interactivity during the training process, the experimental platform is conceived as a multifunctional structure, with several blocks dedicated to specific function through software on

DSP-basis. In fact, we have used DFT and IDFT blocks implemented on a TMS320C30 DSP, and a 32-state Viterbi VLSI decoder implemented on TMS320C55 DSP.

4.2 Simulation results

The simulation results allow us to compare the performances in multipath transmissions.

Fig. 3 shows the packet error-rate for the aforementioned method for a 2-paths (full line) and for a 6-path (dotted line) channel versus the mean signal-noise ratio of a fading channel characterized by γ=5 dB for the reception without diversity and with γ=0 dB for the reception with diversity over two paths.

It is remarkable that in the 2-branch diversity situation, the differences are larger, especially due to the fact that N = 6 requires more past-symbol ISI cancellation. In the same time, it appears that the required SNR for a receiver without diversity is too high to be accepted. The plotted results correspond to the joint blind equalization and detection algorithm using P = 2 and L = 3, considered the optimal values.

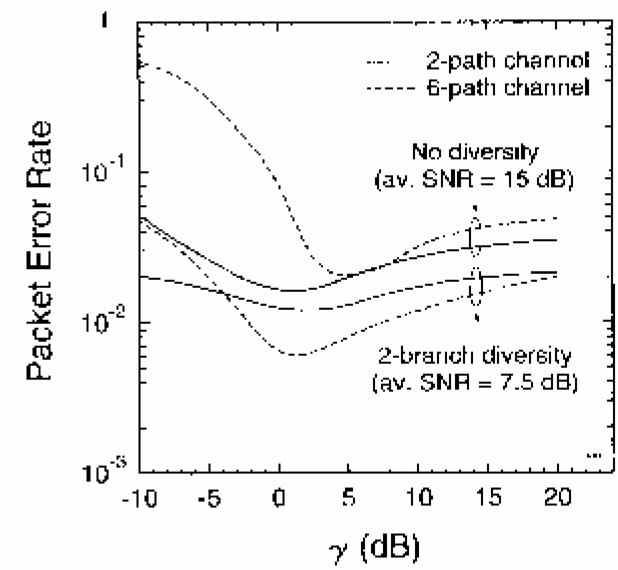

Fig.3. Packet error rate for the proposed communication model

At the first analysis, the proposed scheme gives a coding gain of about 7-8 dB and requires a SNR of only about 10 dB to guarantee a PER of 1%.

5 Conclusions

We can use DSP – based structures in order to implement a BED algorithm that exploit the similarity between OFDM signals detection in frequency-selective channels and the detection of M-PSK signals on time-selective channels. This transmission scheme does not require statistical information about the channel status and consequently no training blocks, allows block-by-block data detection and provides a significant performance improvement over differential detection.

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